A Nonlinear Speed Control for a PM Synchronous Motor Using a Simple Disturbance Estimation Technique

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A nonlinear speed control for a permanent magnet (PM) synchronous motor using a simple disturbance estimation technique is presented. By using a feedback linearization scheme, the nonlinear motor model can be linearized. To compensate an undesirable output performance under the mismatch of the system parameters and load conditions the controller parameters will be estimated by using a disturbance observer theory. Since only the two reduced-order observers are used for the parameter estimation, the observer designs are considerably simple and the computational load of the controller for parameter estimation is negligibly small. The proposed control scheme is implemented on a PM synchronous motor using DSP TMS320C31 and the effectiveness is verified through the comparative experiments.

I. INTRODUCTION

In recent years, feedback linearization techniques have been applied to the control of the nonlinear plants. The basic idea is to first transform a nonlinear system into a linear one by a nonlinear feedback, and then use the well-known linear design techniques to complete the controller design [1]. These techniques, however, require the full knowledge of the system parameters and load conditions with the sufficient accuracy. In general, PM synchronous motor drive systems are faced with unavoidable disturbances or variations of some parameters such as the inertia, viscous friction coefficient and the flux linkage.

This paper presents a nonlinear speed control method for a PM synchronous motor using a simple disturbance estimation technique. Under the assumption that the disturbance torque and flux linkage are unknown, the input-output linearization is performed. The resultant model has the nonlinear disturbances in its input-output relation caused by the unknown disturbance torque and the flux linkage variation. The disturbance torque and flux linkage will be estimated by using a disturbance observer theory. Since only the two reduced-order observers are used for the parameter estimation, the observer designs are considerably simple and the computational load of the controller for parameter estimation is negligibly small. The nonlinear disturbances by the incomplete linearization can be effectively compensated by using the proposed control scheme, and thus, a desired dynamic performance and a zero steady-state error can be obtained. The whole control processing is implemented by the software of DSP TMS320C31 for a PM synchronous motor driven by a three-phase voltage-fed PWM inverter.

II. MODELING OF PM SYNCHRONOUS MOTOR

The stator voltage equations of a PM synchronous motor in the synchronous reference frame are described as follows [2]:

$$v_{qs} = R_s i_{qs} + L_s \dot{i}_{qs} + L_s \omega_r i_{ds} + \lambda_m \omega_r \quad (1)$$

$$v_{ds} = R_s i_{ds} + L_s \dot{i}_{ds} - L_s \omega_r i_{qs} \quad (2)$$

where R_s is the stator resistance, L_s is the stator inductance, ω_r is the electrical rotor angular velocity, and λ_m is the flux linkage established by the permanent magnet. The speed dynamics is expressed as

$$\dot{\omega}_r = \frac{3}{2}\frac{p^2}{J}\lambda_m i_{qs} - \frac{B}{J}\omega_r - \frac{p}{J}T_L \quad (3)$$

where J is the moment of inertia of the rotor and its attached load, B is the viscous friction coefficient, p is the number of pole pairs, and T_L is the load torque. Using ω_r, i_{qs}, and i_{ds} as the state variables, the nonlinear state equation of a PM synchronous motor can be expressed as follows:

$$\dot{x} = f(x) + g_1 v_{qs} + g_2 v_{ds} \quad (4)$$

where $x = [\omega_r \quad i_{qs} \quad i_{ds}]$

$$g_1 = \begin{pmatrix} 0 & \frac{1}{L_s} & 0 \end{pmatrix}^T, \quad g_2 = \begin{pmatrix} 0 & 0 & \frac{1}{L_s} \end{pmatrix}^T$$

$$f(x) = \begin{pmatrix} \frac{3}{2}\frac{p^2}{J}\lambda_m i_{qs} - \frac{B}{J}\omega_r - \frac{p}{J}T_L \\ -\frac{R_s}{L_s} i_{qs} - \omega_r i_{ds} - \frac{\lambda_m}{L_s}\omega_r \\ -\frac{R_s}{L_s} i_{ds} + \omega_r i_{qs} \end{pmatrix}.$$

III. INPUT-OUTPUT FEEDBACK LINEARIZATION

To linearize the nonlinear model in (4), the controlled variable is differentiated with respect to time until the input appears. This can be easily done by introducing the Lie derivative of a state function $h(x): R^n \to R$ along a vector field $f(x)=(f_1,\cdots,f_n)$ as follows [1]:

$$L_f h = \nabla h \cdot f = \sum_{i=1}^{n} \frac{\partial h}{\partial x_i} f_i(x) \quad (5)$$

$$L_f^i h = L_f(L_f^{(i-1)} h) \cdot \quad (6)$$

When the parameters and load of the PM synchronous motor are exactly known, (4) can be transformed to a linear decoupled model of Brunovski canonical form through the linearization technique [3]. Then, the speed controller can be easily designed using the linear control strategy. However, PM synchronous motor drive systems are faced with external disturbances or some parameter variations. This yields a steady-state error as well as a deteriorated transient response due to the incomplete linearization. Since the flux linkage and the disturbance torque caused by the variation of the mechanical parameters have direct influences on the speed control performance, they will be considered as dominant parameters. From the relationship between the developed torque and mechanical load, the torque equation of the machine can be expressed as follows:

$$T_e = J_o\left(\frac{1}{p}\right)\frac{d\omega_r}{dt} + B_o\left(\frac{1}{p}\right)\omega_r + T_d \tag{7}$$

$$T_d = \Delta J\left(\frac{1}{p}\right)\frac{d\omega_r}{dt} + \Delta B\left(\frac{1}{p}\right)\omega_r + T_L \tag{8}$$

where $\Delta J = J - J_o$, $\Delta B = B - B_o$, subscript "o" denotes the nominal value, and T_d is the disturbance torque. Using (7), the speed dynamics is expressed as

$$\dot{\omega}_r = \frac{3}{2}\frac{p^2}{J_o}\lambda_m i_{qs} - \frac{B_o}{J_o}\omega_r - \frac{p}{J_o}T_d. \tag{9}$$

Under the assumption that the disturbance torque and flux linkage are unknown, (4) can be rewritten using the estimated values as follows:

$$\dot{x} = \hat{f}(x) + g_1 v_{qs} + g_2 v_{ds} + d_1 \Delta T_d + d_2 \Delta\lambda_m \tag{10}$$

where $\Delta T_d = T_d - \hat{T}_d$, $\Delta\lambda_m = \lambda_m - \hat{\lambda}_m$

$$d_1 = \begin{pmatrix} -\frac{p}{J_o} & 0 & 0 \end{pmatrix}^T, \quad d_2 = \begin{pmatrix} \frac{3}{2}\frac{p^2}{J_o} i_{qs} & -\frac{\omega_r}{L_s} & 0 \end{pmatrix}^T$$

$$\hat{f}(x) = \begin{pmatrix} \frac{3}{2}\frac{p^2}{J_o}\hat{\lambda}_m i_{qs} - \frac{B_o}{J_o}\omega_r - \frac{p}{J_o}\hat{T}_d \\ -\frac{R_s}{L_s} i_{qs} - \omega_r i_{ds} - \frac{\hat{\lambda}_m}{L_s}\omega_r \\ -\frac{R_s}{L_s} i_{ds} + \omega_r i_{qs} \end{pmatrix}$$

and the symbol "^" denotes the estimated quantity. In order to avoid any zero dynamics, ω_r and i_{ds} are chosen as the outputs [4]. The objective of the control is to maintain the speed and d-axis current to their reference values or trajectories with the desired output dynamic performance. For this objective, the new state variables are defined as follows:

$$z_1 = h_1(x) = \omega_r \tag{11}$$

$$z_2 = L_{\hat{f}} h_1(x) = \frac{3}{2}\frac{p^2}{J_o}\hat{\lambda}_m i_{qs} - \frac{B_o}{J_o}\omega_r - \frac{p}{J_o}\hat{T}_d \tag{12}$$

$$z_3 = h_2(x) = i_{ds} \tag{13}$$

where z_1 is the speed, z_2 is the computed acceleration using the estimated parameters, and z_3 is the d-axis current. By using (11)-(13) as the state variables, (10) can be rewritten as follows:

$$\dot{z}_1 = z_2 + L_{d1}h_1 \cdot \Delta T_d + L_{d2}h_1 \cdot \Delta\lambda_m \tag{14}$$

$$\dot{z}_2 = L_{\hat{f}}^2 h_1 + L_{g1}L_{\hat{f}}h_1 \cdot v_{qs} + \dot{\hat{T}}_d \cdot L_{d1}h_1 + \dot{\hat{\lambda}}_m \cdot L_{d2}h_{1d} + L_{d1}L_{\hat{f}}h_1 \cdot \Delta T + L_{d2}L_{\hat{f}}h_1 \cdot \Delta\lambda_m \tag{15}$$

$$\dot{z}_3 = L_{\hat{f}}h_2 + L_{g2}h_2 \cdot v_{ds} \tag{16}$$

where $L_{d1}h_1 = -\frac{p}{J_o}$, $L_{d2}h_1 = \frac{3}{2}\frac{p^2}{J_o}i_{qs}$

$$L_{g1}L_{\hat{f}}h_1 = \frac{3}{2}\frac{p^2}{J_o}\frac{\hat{\lambda}_m}{L_s}, \quad L_{d1}L_{\hat{f}}h_1 = \frac{pB_o}{J_o^2},$$

$$L_{d2}L_{\hat{f}}h_1 = -\frac{3}{2}\frac{p^2}{J_o}\left(\frac{\hat{\lambda}_m}{L_s}\omega_r + \frac{B_o}{J_o}i_{qs}\right)$$

$$L_{\hat{f}}h_2 = -\frac{R_s}{L_s}i_{ds} + \omega_r i_{qs}, \quad L_{g2} = \frac{1}{L_s}$$

$$L_{\hat{f}}^2 h_1 = \frac{3}{2}\frac{p^2}{J_o}\hat{\lambda}_m\left(-\frac{R_s}{L_s}i_{qs} - \omega_r i_{ds} - \frac{\hat{\lambda}_m}{L_s}\omega_r\right) - \frac{B_o}{J_o}z_2.$$

To linearize the nonlinear state equations in (14)-(16), the control input voltages v_{qs}^* and v_{ds}^* can be expressed as follows:

$$\begin{pmatrix} v_{qs}^* \\ v_{ds}^* \end{pmatrix} = D(x)^{-1}\begin{pmatrix} -L_{\hat{f}}^2 h_1 - \dot{\hat{T}}_d \cdot L_{d1}h_1 - \dot{\hat{\lambda}}_m \cdot L_{d2}h_1 + v_1 \\ -L_{\hat{f}}h_2 + v_2 \end{pmatrix} \tag{17}$$

where v_1 and v_2 are the linear control inputs by which the desired output error dynamics can be assigned, and $D(x)$ is the decoupling matrix defined as

$$D(x) = \begin{pmatrix} L_{g1}L_{\hat{f}}h_1 & 0 \\ 0 & L_{g2}h_2 \end{pmatrix}. \tag{18}$$

To calculate (18), the inverse of $D(x)$ has to be calculated. However, if the estimated value for the flux linkage reaches a particular value, the inverse of $D(x)$ can not be calculated since it becomes a singular matrix. The estimated value that makes $D(x)$ be singular can be obtained using det $D(x)$=0 as $\hat{\lambda}_m = 0$. This singularity has to be considered in the estimation process. Everywhere except for this singular point, the control input voltages in (17) can be always computed. Using (17), the nonlinear motor model becomes an incompletely linearized model expressed as

$$\dot{z}_1 = z_2 + L_{d1}h_1 \cdot \Delta T_d + L_{d2}h_1 \cdot \Delta\lambda_m \tag{19}$$

$$\dot{z}_2 = v_1 + L_{d1}L_{\hat{f}}h_1 \cdot \Delta T_d + L_{d2}L_{\hat{f}}h_1 \cdot \Delta\lambda_m \tag{20}$$

$$\dot{z}_3 = v_2. \tag{21}$$

As a result of the incomplete linearization due to the parameter deviations, the nonlinear model cannot be transformed to a linear decoupled model of Brunovski canonical form and the nonlinear disturbances exist in its input-output relation. Since such disturbances directly influence on the speed control performance, their effects must be quickly removed. Using the transformed states z, the linear control is selected as follows:

$$v_1 = -k_{\omega1}(z_1 - \omega_r^*) - k_{\omega2}(z_2 - \dot{\omega}_r^*) + \ddot{\omega}_r^* \tag{22}$$

$$v_2 = -k_{id}(z_3 - i_{ds}^*) + \dot{i}_{ds}^*. \tag{23}$$

where ω_r^* and i_{ds}^* are the commands for the speed and d-axis current, respectively. Under the perfect parameter matching condition, that is, ΔT_d=0 and $\Delta\lambda_m$=0, this linear control gives the second-order speed error dynamics and the first-order d-axis current error dynamics as follows:

$$(s^2 + k_{\omega2}s + k_{\omega1})e_\omega = se_\omega(0) + \dot{e}_\omega(0) \tag{24}$$

$$(s + k_{id})e_{id} = e_{id}(0) \tag{25}$$

where $e_\omega = \omega_r - \omega_r^*$, $e_{id} = i_{ds} - i_{ds}^*$ and s is a Laplace operator. The desired poles can be easily chosen by adjusting the controller gains $k_{\omega 1}$, $k_{\omega 2}$, and k_{id} through the pole placement technique. Also, to improve the control performance at steady-state, an integral action can be introduced in the linear control as follows:

$$v_{l1} = -k_{\omega I}\int(z_1-\omega_r^*)dt - k_{\omega 1}(z_1-\omega_r^*) - k_{\omega 2}(z_2-\dot{\omega}_r^*) + \ddot{\omega}_r^* \tag{26}$$

$$v_{l2} = -k_{idI}\int(z_3-i_{ds}^*)dt - k_{id}(z_3-i_{ds}^*) + \dot{i}_{ds}^* . \tag{27}$$

IV. PARAMETER ESTIMATION USING DISTURBANCE OBSERVER THEORY

Two disturbance observers are employed for the estimation of the disturbance torque and flux linkage, respectively. To estimate the parameters using the disturbance observer theory, the estimated parameters need to be unknown constant or slowly-varying. Since the flux linkage varies slowly with the temperature rise, λ_m can be assumed to be constant during each sampling interval as follows:

$$\dot{\lambda}_m = 0 . \tag{28}$$

Even though the disturbance torque T_d is not a constant parameter under the mechanical parameter variation such as the inertia and viscous friction coefficient, if the sampling interval is sufficiently fast as compared with the time variation of this unknown disturbance, T_d also can be assumed to be constant during each sampling interval as follows [5]-[6]:

$$\dot{T}_d = 0 . \tag{29}$$

From (28) and the q-axis stator voltage equation in (1), the augmented system for the flux linkage estimation can be expressed as follows:

$$\dot{\boldsymbol{x}}_1 = \boldsymbol{A}\boldsymbol{x}_1 + \boldsymbol{B}u_1 + \boldsymbol{d} \tag{30}$$

$$y_1 = \boldsymbol{C}_1\boldsymbol{x}_1 \tag{31}$$

where $\boldsymbol{x}_1 = [x_{1a} \quad x_{1b}]^T = [i_{qs} \quad \lambda_m]^T$, $\boldsymbol{C}_1 = [1 \quad 0]$,

$u_1 = v_{qs}$, $y_1 = i_{qs}$

$$\boldsymbol{A} = \begin{pmatrix} a_{11} & a_{12} \\ a_{21} & a_{22} \end{pmatrix} = \begin{pmatrix} -\frac{R_s}{L_s} & -\frac{\omega_r}{L_s} \\ 0 & 0 \end{pmatrix},$$

$$\boldsymbol{B} = \begin{pmatrix} b_1 \\ b_2 \end{pmatrix} = \begin{pmatrix} \frac{1}{L_s} \\ 0 \end{pmatrix}, \quad \boldsymbol{d} = \begin{pmatrix} d_1 \\ d_2 \end{pmatrix} = \begin{pmatrix} -\omega_r i_{ds} \\ 0 \end{pmatrix}.$$

The observability matrix for the system in (30) and (31) becomes as follows:

$$\boldsymbol{W}_1 = [\boldsymbol{C}_1^T \quad \boldsymbol{A}^T\boldsymbol{C}_1^T] . \tag{32}$$

Since the rank of a matrix $\boldsymbol{W}_1$ is 2, the full state vector is completely observable. However, the system states are decomposed of the state variable that can be directly measurable and unmeasurable state. To simply estimate this unmeasurable state, a reduced-order observer can be used as follows:

$$\dot{\hat{x}}_{1b} = (a_{22} - L_1a_{12})\hat{x}_{1b} + a_{21}x_{1a} + b_2u_1 + d_2 + L_1(\dot{x}_{1a} - a_{11}x_{1a} - b_1u_1 - d_1) \tag{33}$$

where L_1 is an observer gain for the flux linkage estimation. For the implementation of (33), the derivative of the measured current is required. However, the value of $\dot{x}_{1a}$ is unacceptable because the current signal is generally noisy. To overcome this problem, a new state is defined as

$$x_{c1} = \hat{x}_{1b} - L_1y_1 = \hat{x}_{1b} - L_1x_{1a} . \tag{34}$$

In terms of this state, the observer for the flux linkage estimation is given as follows [19]:

$$\dot{x}_{c1} = (a_{22} - L_1a_{12})\hat{x}_{1b} + (a_{21} - L_1a_{11})x_{1a} + (b_2 - L_1b_1)u_1 + d_2 - L_1d_1 . \tag{35}$$

Then, the derivative of the current signal is no longer used directly. If the estimation error is chosen as $e_1 = (x_{1b} - \hat{x}_{1b})$, the error dynamics of the observer can be expressed as

$$\dot{e}_1 = (a_{22} - L_1a_{12})e_1 . \tag{36}$$

The dynamic characteristics of the observer can be easily determined by the observer gain L_1. From (9) and (29), the augmented system for the disturbance torque estimation can be expressed as follows:

$$\dot{\boldsymbol{x}}_2 = \boldsymbol{G}\boldsymbol{x}_2 + \boldsymbol{H}u_2 \tag{37}$$

$$y_2 = \boldsymbol{C}_2\boldsymbol{x}_2 \tag{38}$$

where $\boldsymbol{x}_2 = [x_{2a} \quad x_{2b}]^T = [\omega_r \quad T_d]^T$, $\boldsymbol{C}_2 = [1 \quad 0]$,

$u_2 = i_{qs}$ $y_2 = \omega_r$

$$\boldsymbol{G} = \begin{pmatrix} g_{11} & g_{12} \\ g_{21} & g_{22} \end{pmatrix} = \begin{pmatrix} -\frac{B_o}{J_o} & -\frac{p}{J_o} \\ 0 & 0 \end{pmatrix}, \quad \boldsymbol{H} = \begin{pmatrix} h_1 \\ h_2 \end{pmatrix} = \begin{pmatrix} \frac{3}{2}\frac{p^2}{J_o}\lambda_m \\ 0 \end{pmatrix}.$$

Since the rank of the observability matrix for the system in (37) and (38)

$$\boldsymbol{W}_2 = [\boldsymbol{C}_2^T \quad \boldsymbol{G}^T\boldsymbol{C}_2^T] \tag{39}$$

is 2, the full state vector is completely observable. Similarly, the system states are decomposed of the directly measurable and unmeasurable states. A reduced-order observer for estimating this unmeasurable state can be expressed as follows:

$$x_{c2} = \hat{x}_{2b} - L_2y_2 = \hat{x}_{2b} - L_2x_{2a} \tag{40}$$

$$\dot{x}_{c2} = (g_{22} - L_2g_{12})\hat{x}_{2b} + (g_{21} - L_2g_{11})x_{2a} + (h_2 - L_2\hat{h}_1)u_2 \tag{41}$$

where $\hat{h}_1 = \frac{3}{2}\frac{p^2}{J_o}\hat{\lambda}_m$ and L_2 is an observer gain for the disturbance torque estimation. Notice that the estimated value $\hat{\lambda}_m$ is used in (41). If the estimation error is chosen as $e_2 = (x_{2b} - \hat{x}_{2b})$, the error dynamics of the observer can be expressed as

$$\dot{e}_2 = (g_{22} - L_2g_{12})e_2 - \Delta h_1 \cdot L_2u_2 \tag{42}$$

where $\Delta h_1 = h_1 - \hat{h}_1 = \frac{3}{2}\frac{p^2}{J_o}\Delta\lambda_m$.

Because the term Δh_1 disappears to zero as the estimated flux linkage converges to its real value, the dynamic characteristics of the observer can be determined by the observer gain L_2. The flux linkage and disturbance torque can be simultaneously estimated by using (34)-(35) and (40)-(41). The estimated parameters are used for the computation of the control input voltages in (17) and the state z_2 in (12) to compensate the nonlinear disturbances caused by the parameter variations.

V. EXPERIMENTS

The overall block diagram for the proposed control scheme is shown in Fig. 1. To improve the control performance at steady-state, the linear control with an integrator in (26) and (27) is employed. Then, the large portions of the unknown disturbance are compensated by the disturbance observer and the remaining small portions are effectively rejected by the integral action. The nominal parameters of a PM synchronous motor used for the simulations and experiments are listed in Table I. For the performance comparison, four control methods including the proposed control scheme will be used, which are listed in Table II. The d-axis current command is given as zero and the speed trajectory command is given as follows:

$$\omega_r^* = \frac{\omega_{rf}}{T_f}t - \frac{\omega_{rf}}{2\pi}\sin\left(\frac{2\pi t}{T_f}\right) \quad (46)$$

$$\dot{\omega}_r^* = \frac{\omega_{rf}}{T_f} - \frac{\omega_{rf}}{T_f}\cos\left(\frac{2\pi t}{T_f}\right) \quad (47)$$

$$\ddot{\omega}_r^* = 2\pi\frac{\omega_{rf}}{T_f^2}\sin\left(\frac{2\pi t}{T_f}\right) \quad (48)$$

where ω_{rf} is the desired final speed and T_f is the time when the speed command reaches from zero to ω_{rf}.

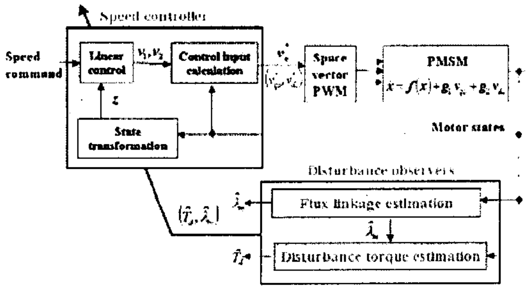


Fig.1 Overall block diagram for proposed control scheme

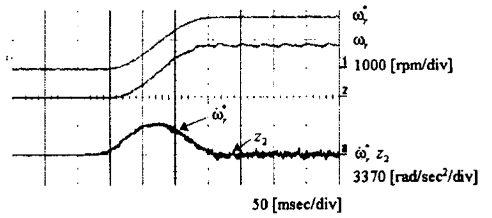
Figs. 2 (a) and (b) show the experimental results for the method A under the nominal parameters and under the inertia variations, respectively. The gains for the linear control are selected as $k_{\omega 2}$ =400, $k_{\omega 1}$ =80000, and k_{id}=1000 so that the poles of the speed error dynamics and the d-axis current error dynamics are determined as $-200\pm j200$ and -1000, respectively. It is shown in Fig. 2 (a) that the speed and acceleration commands can be well tracked. However, as can be shown in Fig. 2 (b), this control scheme gives an undesirable dynamic performance under the inertia variations. Also, the computed acceleration z_2 shows the large transient errors. Figs. 3 (a) and (b) show the experimental results for the method B. Fig. 3 (a) shows the case for the inertia variations, and Fig. 3 (b) shows the case for both the flux linkage and inertia variations, respectively. The speed responses in Fig. 3 (a) show a good dynamic performance under the inertia variations since the effective disturbance torque can be estimated by the disturbance torque observer. However, under the flux linkage variation, there exists an estimation error in the disturbance torque. This yields a steady-state error in the speed response as can be shown in Fig. 3 (b).

TABLE I
SPECIFICATIONS OF A PM SYNCHRONOUS MOTOR

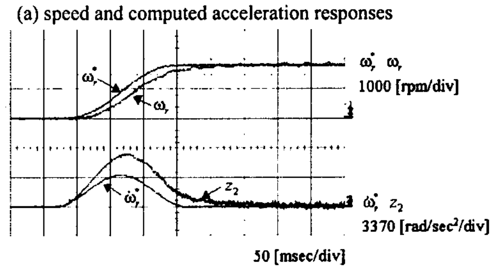
Rated power	400 W	Number of poles	4
Rated torque	1.27 Nm	Stator resistance	3.0 Ω
Magnetic flux	0.153 Wb	Stator inductance	10.5 mH
Rated speed	3000 rpm	Moment of inertia	1.75×10^{-4} Nm·s^2

TABLE II
FOUR CONTROL METHODS FOR THE PERFORMANCE COMPARISON

Method A	Linear control in (22) and (23) without disturbance torque observer
Method B	Linear control in (22) and (23) with disturbance torque observer
Method C	Linear control in (26) and (27) with disturbance torque observer
Proposed	Linear control in (26) and (27) with torque and flux observer

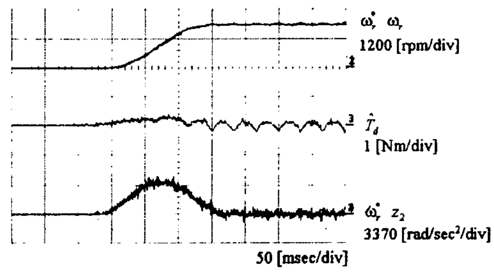


(a) speed and computed acceleration responses

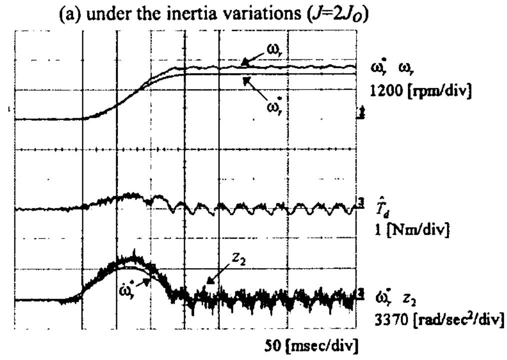


(b) speed transient response and computed acceleration under J=2J_O

Fig.2 Experimental results for method A



(a) under the inertia variations (J=2J_O)



(b) under the parameter variations (J=2J_O and $\Delta\lambda_m$= -0.2λ_{mo})

Fig.3 Experimental results of method B

As mentioned earlier, this steady-state speed error can be removed by introducing the integral control and disturbance torque observer. The integral gains are chosen as $k_{\omega I}$=2000000 and k_{idI}=500000. However, the speed response generally shows a large transient with a fixed integral gain. The speed responses for the method C are shown in Fig. 4. Figs. 5 (a) and (b) shows the experimental results of the proposed control scheme. Fig. 5 shows the speed transient response and parameter estimation performance when the inertia is varied to $2J_o$ and $\Delta\lambda_m$ $(\lambda_m - \hat{\lambda}_m)$ is -20% of its nominal value. Even under both the flux linkage and inertia variations, the speed response of the proposed control scheme is unaffected by these variations and shows a good dynamic performance. The current waveforms and computed acceleration response are shown in Fig. 5 (b). It can be shown that the *d*-axis current is well regulated to zero. The computed acceleration z_2 using the estimated parameters initially shows a large transient error due to the parameter variations. However, as the disturbance torque and flux linkage are estimated, z_2 is well controlled to the acceleration command.

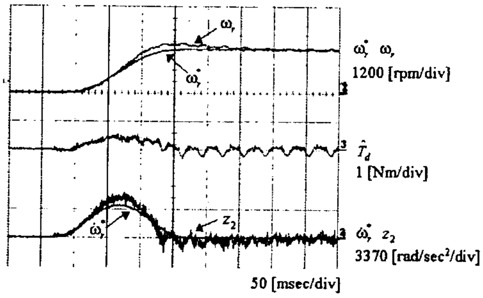
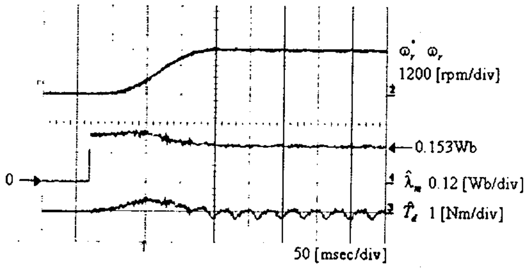
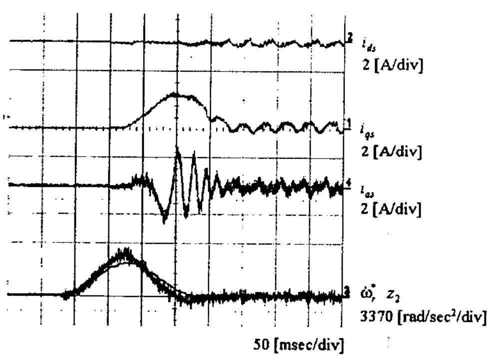


Fig. 4 Experimental results of method C (J=$2J_o$ and $\Delta\lambda_m$= $-0.2\lambda_{mo}$)



(a) speed transient response and parameter estimation performance



(b) Current and acceleration waveforms

Fig. 5 Experimental results of the proposed control scheme under J=$2J_o$ and $\Delta\lambda_m$= $-0.2\lambda_{mo}$

VI. CONCLUSIONS

A nonlinear speed control method for a PM synchronous motor using a simple disturbance estimation technique has been proposed for the design of the speed tracking controller. By using this, a systematic design approach for a speed controller can be accomplished without considering a separate inner-loop current regulator. Through the various comparative experimental results, it is verified that the proposed control scheme yields a robust control performance even under the presence of the parameter variation and the external disturbances caused by the inertia and load changes. Since only the two reduced-order observers are used for the parameter estimation, the observer designs are considerably simple and the computational load of the controller for parameter estimation is negligibly small as compared with the adaptation technique in [3].. As a result, without requiring a complex controller design such as the adaptation technique an improved robustness against the parameter variations can be obtained. Thus, a speed response gives a desired dynamic performance and a zero steady-state error, which is not affected by the load torque disturbance and the variation of the motor and mechanical parameters. The whole control system is realized using DSP TMS320C31 for a PM synchronous motor driven by a three-phase voltage-fed PWM inverter.

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